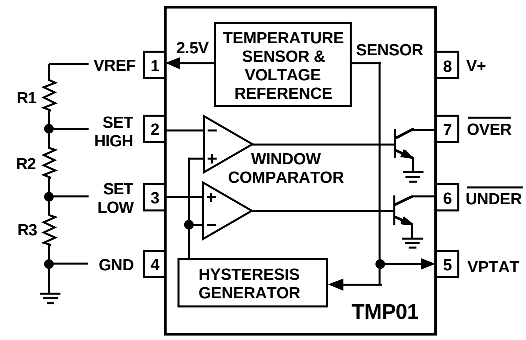
Low Power, Programmable Temperature Controller

TMP01*

FEATURES
- **–55°C to +125°C (–67°F to +257°F) Operation**
- **±1.0°C Accuracy Over Temperature (typ)**
- **Temperature-Proportional Voltage Output**
- **User Programmable Temperature Trip Points**
- **User Programmable Hysteresis**
- **20 mA Open Collector Trip Point Outputs**
- **TTL/CMOS Compatible**
- **Single-Supply Operation (4.5 V to 13.2 V)**
- **Low Cost 8-Pin DIP and SO Packages**

APPLICATIONS
- **Over/Under Temperature Sensor and Alarm**
- **Board Level Temperature Sensing**
- **Temperature Controllers**
- **Electronic Thermostats**
- **Thermal Protection**
- **HVAC Systems**
- **Industrial Process Control**
- **Remote Sensors**

FUNCTIONAL BLOCK DIAGRAM



GENERAL DESCRIPTION

The TMP01 is a temperature sensor which generates a voltage output proportional to absolute temperature and a control signal from one of two outputs when the device is either above or below a specific temperature range. Both the high/low temperature trip points and hysteresis (overshoot) band are determined by user-selected external resistors. For high volume production, these resistors are available on-board.

The TMP01 consists of a bandgap voltage reference combined with a pair of matched comparators. The reference provides both a constant 2.5 V output and a voltage proportional to absolute temperature (VPTAT) which has a precise temperature coefficient of 5 mV/K and is 1.49 V (nominal) at +25°C. The comparators compare VPTAT with the externally set temperature trip points and generate an open-collector output signal when one of their respective thresholds has been exceeded.

Hysteresis is also programmed by the external resistor chain and is determined by the total current drawn out of the 2.5 V reference. This current is mirrored and used to generate a hysteresis offset voltage of the appropriate polarity after a comparator has been tripped. The comparators are connected in parallel, which guarantees that there is no hysteresis overlap and eliminates erratic transitions between adjacent trip zones.

The TMP01 utilizes proprietary thin-film resistors in conjunction with production laser trimming to maintain a temperature accuracy of ±1°C (typ) over the rated temperature range, with excellent linearity. The open-collector outputs are capable of sinking 20 mA, enabling the TMP01 to drive control relays directly. Operating from a +5 V supply, quiescent current is only 500 µA (max).

The TMP01 is available in the low cost 8-pin epoxy mini-DIP and SO (small outline) packages, and in die form.

***Protected by U.S. Patent No. 5,195,827.**

REV. C

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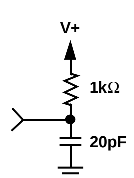
(V+ = +5 V, GND = 0 V, –40°C ≤ T_A ≤ +85°C unless otherwise noted)

Parameter	Symbol	Conditions	Min	Typ	Max	Units
INPUTS SET HIGH, SET LOW						
Offset Voltage	V_{OS}			0.25		mV
Offset Voltage Drift	TCV_{OS}			3		µV/°C
Input Bias Current, "E"	I_B			25	50	nA
Input Bias Current, "F"	I_B			25	100	nA
OUTPUT VPTAT[1]						
Output Voltage	VPTAT	T_A = +25°C, No Load		1.49		V
Scale Factor	TC_{VPTAT}			5		mV/K
Temperature Accuracy, "E"		T_A = +25°C, No Load	–1.5	±0.5	1.5	°C
Temperature Accuracy, "F"		T_A = +25°C, No Load	–3	±1.0	3	°C
Temperature Accuracy, "E"		10°C < T_A < 40°C, No Load		±0.75		°C
Temperature Accuracy, "F"		10°C < T_A < 40°C, No Load		±1.5		°C
Temperature Accuracy, "E"		–40°C < T_A < 85°C, No Load	–3.0	±1	3.0	°C
Temperature Accuracy, "F"		–40°C < T_A < 85°C, No Load	–5.0	±2	5.0	°C
Temperature Accuracy, "E"		–55°C < T_A < 125°C, No Load		±1.5		°C
Temperature Accuracy, "F"		–55°C < T_A < 125°C, No Load		±2.5		°C
Repeatability Error	ΔVPTAT	Note 4		0.25		Degree
Long Term Drift Error		Notes 2 and 6		0.25	0.5	Degree
Power Supply Rejection Ratio	PSRR	T_A = +25°C, 4.5 V ≤ V+ ≤ 13.2 V		±0.02	±0.1	%/V
OUTPUT VREF						
Output Voltage, "E"	VREF	T_A = +25°C, No Load	2.495	2.500	2.505	V
Output Voltage, "F"	VREF	T_A = +25°C, No Load	2.490	2.500	2.510	V
Output Voltage, "E"	VREF	–40°C < T_A < 85°C, No Load	2.490	2.500	2.510	V
Output Voltage, "F"	VREF	–40°C < T_A < 85°C, No Load	2.485	2.500	2.515	V
Output Voltage, "E"	VREF	–55°C < T_A < 125°C, No Load		2.5 ± 0.01		V
Output Voltage, "F"	VREF	–55°C < T_A < 125°C, No Load		2.5 ± 0.015		V
Drift	TC_{VREF}			–10		ppm/°C
Line Regulation		4.5 V ≤ V+ ≤ 13.2 V		±0.01	±0.05	%/V
Load Regulation		10 µA ≤ I_{VREF} ≤ 500 µA		±0.1	±0.25	%/mA
Output Current, Zero Hysteresis	I_{VREF}			7		µA
Hysteresis Current Scale Factor	SF_{HYS}	(Note 1)		5.0		µA/°C
Turn-On Settling Time		To Rated Accuracy		25		µs
OPEN-COLLECTOR OUTPUTS OVER, UNDER						
Output Low Voltage	V_{OL}	I_{SINK} = 1.6 mA		0.25	0.4	V
Output Low Voltage	V_{OL}	I_{SINK} = 20 mA		0.6		V
Output Leakage Current	I_{OH}	V+ = 12 V		1	100	µA
Fall Time	t_{HL}	See Test Load		40		ns
POWER SUPPLY						
Supply Range	V+		4.5		13.2	V
Supply Current	I_{SY}	Unloaded, +V = 5 V		400	500	µA
Supply Current	I_{SY}	Unloaded, +V = 13.2 V		450	800	µA
Power Dissipation	P_{DISS}	+V = 5 V		2.0	2.5	mW

NOTES
[1]K = °C + 273.15.
[2]Guaranteed but not tested.
[3]Does not consider errors caused by heating due to dissipation of output load currents.
[4]Maximum deviation between +25°C readings after temperature cycling between –55°C and +125°C.
[5]Typical values indicate performance measured at T_A = +25°C.
[6]Observed in a group sample over an accelerated life test of 500 hours at 150°C.

Specifications subject to change without notice.

Test Load



TMP01FJ–SPECIFICATIONS

TO-99 Metal Can Package (V+ = +5 V, GND = 0 V, –40°C ≤ T_A ≤ +85°C unless otherwise noted)

Parameter	Symbol	Conditions	Min	Typ	Max	Units
INPUTS SET HIGH, SET LOW						
Offset Voltage	V_{OS}			0.25		mV
Offset Voltage Drift	TCV_{OS}			3		µV/°C
Input Bias Current, "F"	I_B			25	100	nA
OUTPUT VPTAT[1]						
Output Voltage	VPTAT	T_A = +25°C, No Load		1.49		V
Scale Factor	TC_{VPTAT}			5		mV/K
Temperature Accuracy, "F"		T_A = +25°C, No Load	–3	±1.0	3	°C
Temperature Accuracy, "F"		10°C < T_A < 40°C, No Load		±1.5		°C
Temperature Accuracy, "F"		–40°C < T_A < 85°C, No Load	–5.0	±2	5.0	°C
Temperature Accuracy, "F"		–55°C < T_A < 125°C, No Load		±2.5		°C
Repeatability Error	ΔVPTAT	Note 4		0.25		Degree
Long Term Drift Error		Notes 2 and 6		0.25	0.5	Degree
Power Supply Rejection Ratio	PSRR	T_A = +25°C, 4.5 V ≤ V+ ≤ 13.2 V		±0.02	±0.1	%/V
OUTPUT VREF						
Output Voltage, "F"	VREF	T_A = +25°C, No Load	2.490	2.500	2.510	V
Output Voltage, "F"	VREF	–40°C < T_A < 85°C, No Load	2.480	2.500	2.520	V
Output Voltage, "F"	VREF	–55°C < T_A < 125°C, No Load		2.5 ± 0.015		V
Drift	TC_{VREF}			–10		ppm/°C
Line Regulation		4.5 V ≤ V+ ≤ 13.2 V		±0.01	±0.05	%/V
Load Regulation		10 µA ≤ I_{VREF} ≤ 500 µA		±0.1	±0.25	%/mA
Output Current, Zero Hysteresis	I_{VREF}			7		µA
Hysteresis Current Scale Factor	SF_{HYS}	(Note 1)		5.0		µA/°C
Turn-On Settling Time		To Rated Accuracy		25		µs
OPEN-COLLECTOR OUTPUTS OVER, UNDER						
Output Low Voltage	V_{OL}	I_{SINK} = 1.6 mA		0.25	0.4	V
Output Low Voltage	V_{OL}	I_{SINK} = 20 mA		0.6		V
Output Leakage Current	I_{OH}	V+ = 12 V		1	100	µA
Fall Time	t_{HL}	See Test Load, Note 2		40		ns
POWER SUPPLY						
Supply Range	V+		4.5		13.2	V
Supply Current	I_{SY}	Unloaded, +V = 5 V		400	500	µA
Supply Current	I_{SY}	Unloaded, +V = 13.2 V		450	800	µA
Power Dissipation	P_{DISS}	+V = 5 V		2.0	2.5	mW

NOTES

[1]K = °C + 273.15.

[2]Guaranteed but not tested.

[3]Does not consider errors caused by heating due to dissipation of output load currents.

[4]Maximum deviation between +25°C readings after temperature cycling between –55°C and +125°C.

[5]Typical values indicate performance measured at T_A = +25°C.

[6]Observed in a group sample over an accelerated life test of 500 hours at 150°C.

Specifications subject to change without notice.

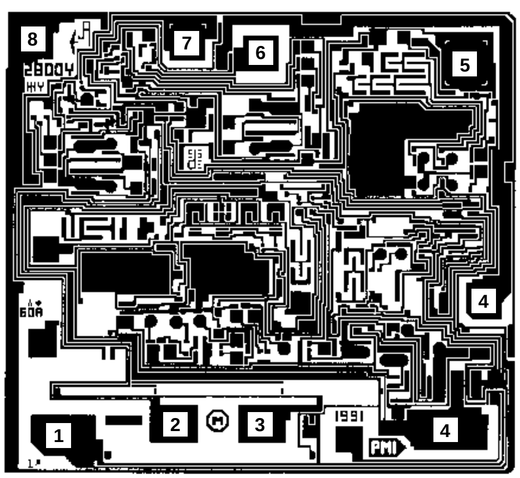
WAFER TEST LIMITS (V_{DD} = +5.0 V, GND = 0 V, T_A = +25°C, unless otherwise noted)

Parameter	Symbol	Conditions	Min	Typ	Max	Units
INPUTS SET HIGH, SET LOW						
Input Bias Current	I_B				100	nA
OUTPUT VPTAT						
Temperature Accuracy		T_A = +25°C, No Load			1.5	°C
OUTPUT VREF						
Nominal Value	VREF	T_A = +25°C, No Load	2.490		2.510	V
Line Regulation		4.5 V ≤ V+ ≤ 13.2 V			±0.05	%/V
Load Regulation		10 µA ≤ I_{VREF} ≤ 500 µA			±0.25	%/mA
OPEN-COLLECTOR OUTPUTS OVER, UNDER						
Output Low Voltage	V_{OL}	I_{SINK} = 1.6 mA			0.4	mV
Output Low Voltage	V_{OL}	I_{SINK} = 20 mA			1.0	V
Output Leakage Current	I_{OH}				100	µA
POWER SUPPLY						
Supply Range	V+		4.5		13.2	V
Supply Current	I_{SY}	Unloaded			600	µA

NOTES
Electrical tests are performed at wafer probe to the limits shown. Due to variations in assembly methods and normal yield loss, yield after packaging is not guaranteed for standard product dice. Consult factory to negotiate specifications based on dice lot qualification through sample lot assembly and testing.

DICE CHARACTERISTICS

Die Size 0.078 × 0.071 inch, 5,538 sq. mils
(1.98 × 1.80 mm, 3.57 sq. mm)
Transistor Count: 105



1. VREF
2. SETHIGH
3. SETLOW
4. GND (TWO PLACES) (CONNECTED TO SUBSTRATE)
5. VPTAT
6. UNDER
7. OVER
8. V+

For additional DICE ordering information, refer to databook.

ABSOLUTE MAXIMUM RATINGS

Maximum Supply Voltage . . . . –0.3 V to +15 V
Maximum Input Voltage
(SETHIGH, SETLOW) . . . .–0.3 V to [(V+) +0.3 V]
Maximum Output Current (VREF, VPTAT) . . . . 2 mA
Maximum Output Current (Open Collector Outputs) . . 50 mA
Maximum Output Voltage (Open Collector Outputs) . . . . 15 V
Operating Temperature Range . . . .–55°C to +150°C
Dice Junction Temperature . . . . +150°C
Storage Temperature Range . . . . – 65°C to +150°C
Lead Temperature (Soldering, 60 sec) . . . . +300°C

Package Type	θ_{JA}	θ_{JC}	Units
8-Pin Plastic DIP (P)	103[1]	43	°C/W
8-Lead SOIC (S)	158[2]	43	°C/W
8-Lead TO-99 Can (J)	150[1]	18	°C/W

NOTES
[1]θ_{JA} is specified for device in socket (worst case conditions).
[2]θ_{JA} is specified for device mounted on PCB.

CAUTION

1. Stresses above those listed under "Absolute Maximum Ratings" may cause permanent damage to the device. This is a stress rating only and functional operation at or above this specification is not implied. Exposure to the above maximum rating conditions for extended periods may affect device reliability.
2. Digital inputs and outputs are protected, however, permanent damage may occur on unprotected units from high energy electrostatic fields. Keep units in conductive foam or packaging at all times until ready to use. Use proper antistatic handling procedures.
3. Remove power before inserting or removing units from their sockets.

ORDERING GUIDE

Model/Grade	Temperature Range[1]	Package Description	Package Option
TMP01EP	XIND	Plastic DIP	N-8
TMP01FP	XIND	Plastic DIP	N-8
TMP01ES	XIND	SOIC	SO-8
TMP01FS	XIND	SOIC	SO-8
TMP01FJ[2]	XIND	TO-99 Can	H-08A
TMP01GBC	+25°C	Die	

NOTES
[1]XIND = –40°C to +85°C.
[2]Consult factory for availability of MIL/883 version in TO-99 can.

GENERAL DESCRIPTION

The TMP01 is a very linear voltage-output temperature sensor, with a window comparator that can be programmed by the user to activate one of two open-collector outputs when a predetermined temperature setpoint voltage has been exceeded. A low drift voltage reference is available for setpoint programming.

The temperature sensor is basically a very accurately temperature compensated, bandgap-type voltage reference with a buffered output voltage proportional to absolute temperature (VPTAT), accurately trimmed to a scale factor of 5 mV/K. See the Applications Information following.

The low drift 2.5 V reference output VREF is easily divided externally with fixed resistors or potentiometers to accurately establish the programmed heat/cool setpoints, independent of temperature. Alternatively, the setpoint voltages can be supplied by other ground referenced voltage sources such as user-programmed DACs or controllers. The high and low setpoint voltages are compared to the temperature sensor voltage, thus creating a two-temperature thermostat function. In addition, the total output current of the reference (I_{VREF}) determines the magnitude of the temperature hysteresis band. The open collector outputs of the comparators can be used to control a wide variety of devices.

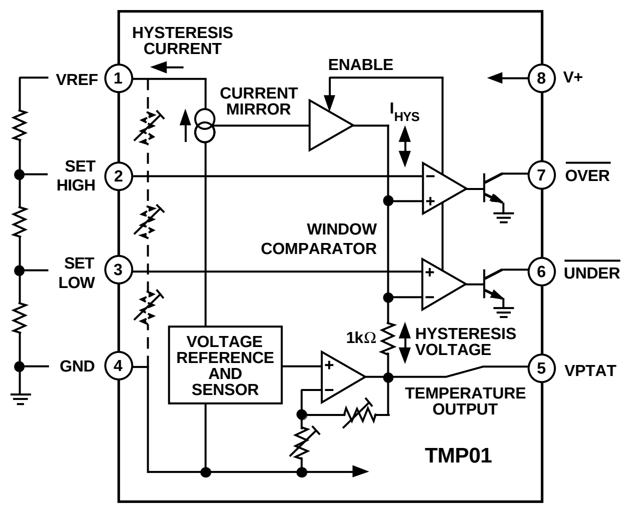


Figure 1. Detailed Block Diagram

Temperature Hysteresis

The temperature hysteresis is the number of degrees beyond the original setpoint temperature that must be sensed by the TMP01 before the setpoint comparator will be reset and the output disabled. Figure 2 shows the hysteresis profile. The hysteresis is programmed by the user by setting a specific load on the reference voltage output VREF. This output current I_{VREF} is also called the hysteresis current, which is mirrored internally and fed to a buffer with an analog switch.

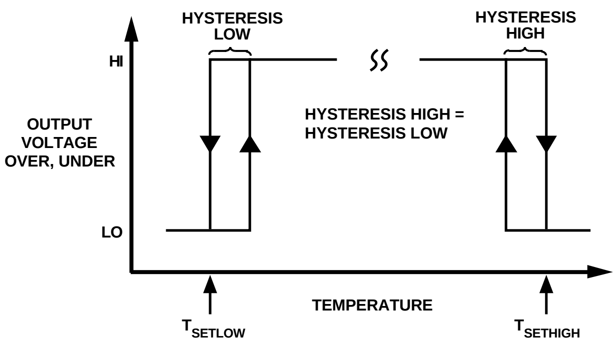


Figure 2. TMP01 Hysteresis Profile

After a temperature setpoint has been exceeded and a comparator tripped, the buffer output is enabled. The output is a current of the appropriate polarity which generates a hysteresis offset voltage across an internal 1000 Ω resistor at the comparator input. The comparator output remains “on” until the voltage at the comparator input, now equal to the temperature sensor voltage VPTAT summed with the hysteresis offset, has returned to the programmed setpoint voltage. The comparator then returns LOW, deactivating the open-collector output and disabling the hysteresis current buffer output. The scale factor for the programmed hysteresis current is:

$$I_{HYS} = I_{VREF} = 5\ \mu A/°C + 7\ \mu A$$

Thus since VREF = 2.5 V, with a reference load resistance of 357 kΩ or greater (output current 7 µA or less), the temperature setpoint hysteresis will be zero degrees. See the temperature programming discussion below. Larger values of load resistance will only decrease the output current below 7 µA and will have no effect on the operation of the device. The amount of hysteresis is determined by selecting a value of load resistance for VREF, as shown below.

Programming the TMP01

In the basic fixed-setpoint application utilizing a simple resistor ladder voltage divider, the desired temperature setpoints are programmed in the following sequence:

1. Select the desired hysteresis temperature.
2. Calculate the hysteresis current I_{VREF}.
3. Select the desired setpoint temperatures.
4. Calculate the individual resistor divider ladder values needed to develop the desired comparator setpoint voltages at SETHIGH and SETLOW.

The hysteresis current is readily calculated, as shown. For example, for 2 degrees of hysteresis, I_{VREF} = 17 µA. Next, the setpoint voltages $V_{SETHIGH}$ and V_{SETLOW} are determined using the VPTAT scale factor of 5 mV/K = 5 mV/(°C + 273.15), which is 1.49 V for +25°C. We then calculate the divider resistors, based on those setpoints. The equations used to calculate the resistors are:

$$V_{SETHIGH} = (T_{SETHIGH} + 273.15)(5\ mV/°C)$$

$$V_{SETLOW} = (T_{SETLOW} + 273.15)\ (5\ mV/°C)$$

$$R1\ (k\Omega) = (V_{VREF} - V_{SETHIGH})/I_{VREF} = (2.5\ V - V_{SETHIGH})/I_{VREF}$$

$$R2\ (k\Omega) = (V_{SETHIGH} - V_{SETLOW})/I_{VREF}$$

$$R3\ (k\Omega) = V_{SETLOW}/I_{VREF}$$

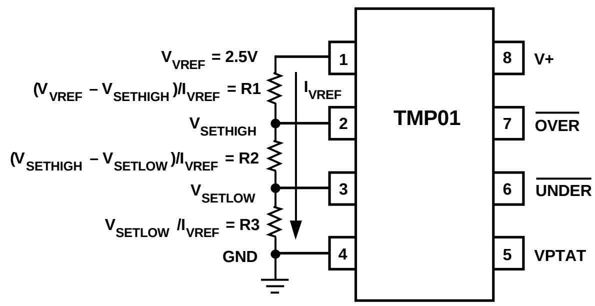


Figure 3. TMP01 Setpoint Programming

The total R1 + R2 + R3 is equal to the load resistance needed to draw the desired hysteresis current from the reference, or I_{VREF}.

The formulas shown above are also helpful in understanding the calculation of temperature setpoint voltages in circuits other than the standard two-temperature thermostat. If a setpoint function is not needed, the appropriate comparator should be disabled. SETHIGH can be disabled by tying it to V+, SETLOW by tying it to GND. Either output can be left unconnected.

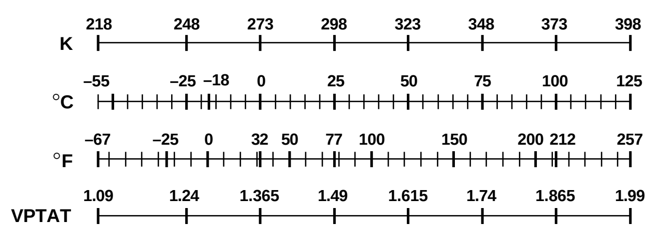


Figure 4. Temperature—VPTAT Scale

Understanding Error Sources

The accuracy of the VPTAT sensor output is well characterized and specified, however preserving this accuracy in a heating or cooling control system requires some attention to minimizing the various potential error sources. The internal sources of setpoint programming error include the initial tolerances and temperature drifts of the reference voltage VREF, the setpoint comparator input offset voltage and bias current, and the hysteresis current scale factor. When evaluating setpoint programming errors, remember that any VREF error contribution at the comparator inputs is reduced by the resistor divider ratios. The comparator input bias current (inputs SETHIGH, SETLOW) drops to less than 1 nA (typ) when the comparator is tripped. This can account for some setpoint voltage error, equal to the change in bias current times the effective setpoint divider ladder resistance to ground.

The thermal mass of the TMP01 package and the degree of thermal coupling to the surrounding circuitry are the largest factors in determining the rate of thermal settling, which ultimately determines the rate at which the desired temperature measurement accuracy may be reached. Thus, one must allow sufficient time for the device to reach the final temperature. The typical thermal time constant for the plastic package is approximately 140 seconds in still air! Therefore, to reach the final temperature accuracy within 1%, for a temperature change of 60 degrees, a settling time of 5 time constants, or 12 minutes, is necessary.

The setpoint comparator input offset voltage and zero hysteresis current affect setpoint error. While the 7 μA zero hysteresis current allows the user to program the TMP01 with moderate resistor divider values, it does vary somewhat from device to device, causing slight variations in the actual hysteresis obtained in practice. Comparator input offset directly impacts the programmed setpoint voltage and thus the resulting hysteresis band, and must be included in error calculations.

External error sources to consider are the accuracy of the programming resistors, grounding error voltages, and the overall problem of thermal gradients. The accuracy of the external programming resistors directly impacts the resulting setpoint accuracy. Thus in fixed-temperature applications the user should select resistor tolerances appropriate to the desired programming accuracy. Resistor temperature drift must be taken into account also. This effect can be minimized by selecting good quality components, and by keeping all components in close thermal proximity. Applications requiring high measurement accuracy require great attention to detail regarding thermal gradients. Careful circuit board layout, component placement, and protection from stray air currents are necessary to minimize common thermal error sources.

Also, the user should take care to keep the bottom of the setpoint programming divider ladder as close to GND (Pin 4) as possible to minimize errors due to IR voltage drops and coupling of external noise sources. In any case, a 0.1 μF capacitor for power supply bypassing is always recommended at the chip.

Safety Considerations In Heating And Cooling System Design

Designers should anticipate potential system fault conditions which may result in significant safety hazards which are outside the control of and cannot be corrected by the TMP01-based circuit. Governmental and industrial regulations regarding safety requirements and standards for such designs should be observed where applicable.

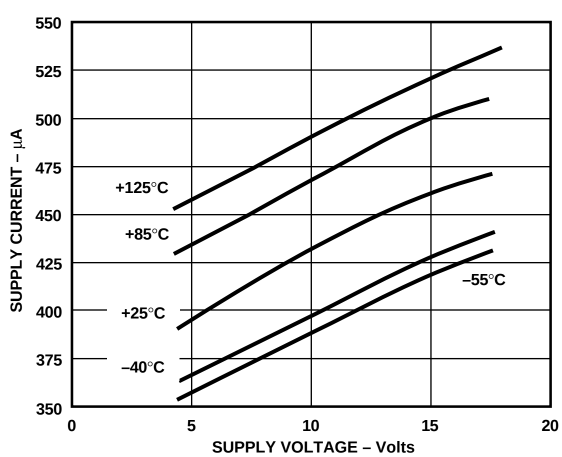


Figure 5. Supply Current vs. Supply Voltage

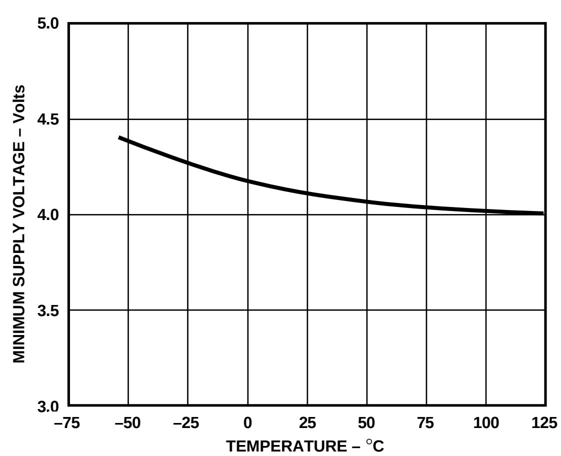


Figure 6. Minimum Supply Voltage vs. Temperature

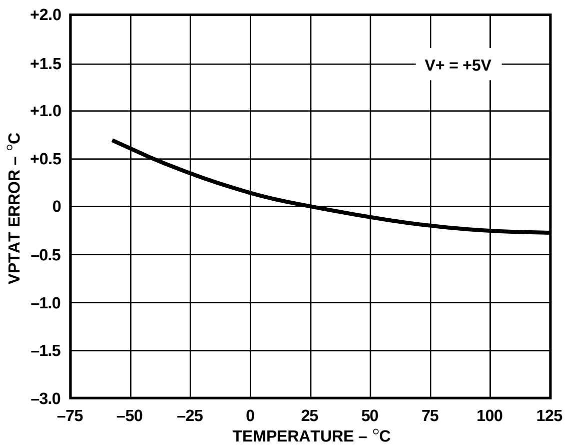


Figure 7. VPTAT Accuracy vs. Temperature

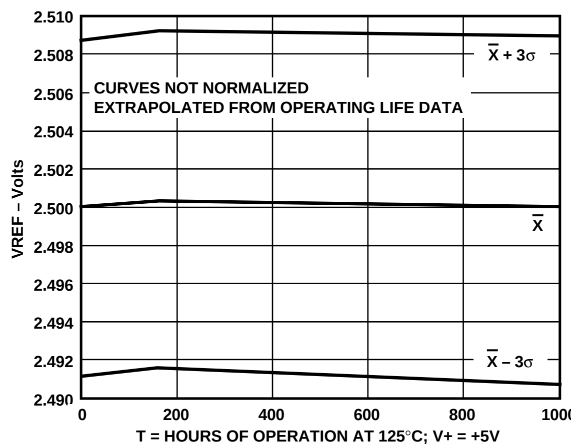


Figure 10. VREF Long Term Drift Accelerated by Burn-In

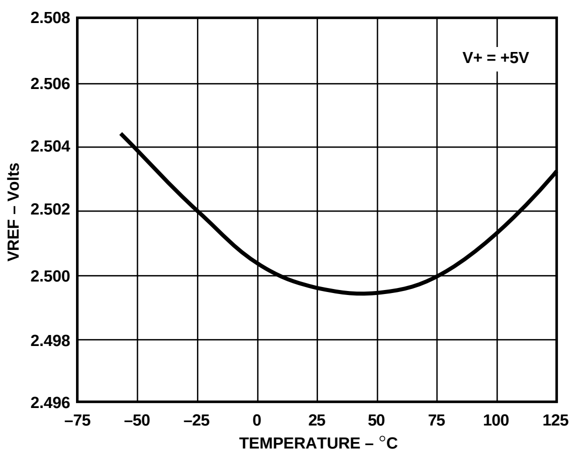


Figure 8. VREF Accuracy vs. Temperature

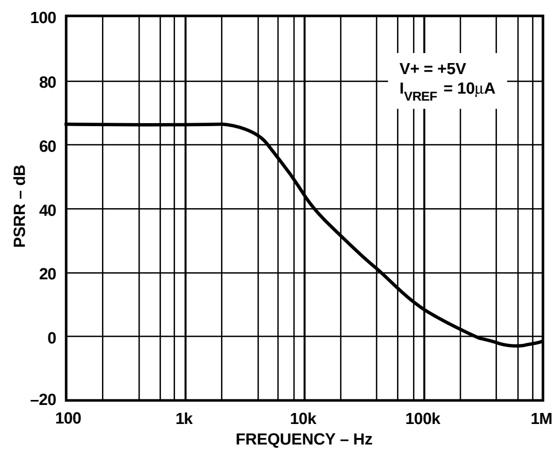


Figure 11. VREF Power Supply Rejection vs. Frequency

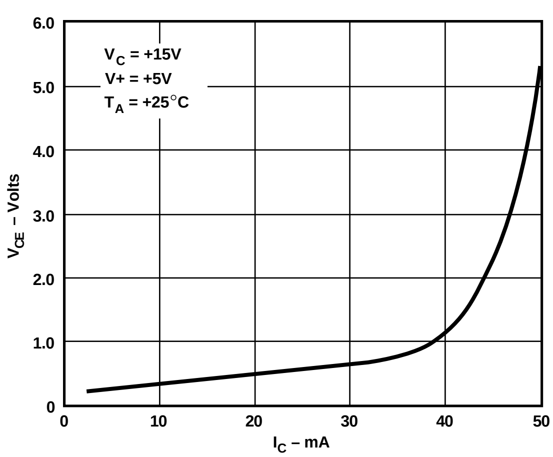


Figure 9. Open-Collector Output ($\overline{\text{OVER}}$, $\overline{\text{UNDER}}$) Saturation Voltage vs. Output Current

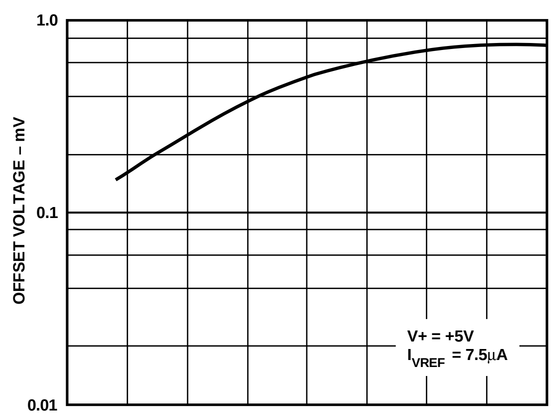


Figure 12. Set High, Set Low Input Offset Voltage vs. Temperature

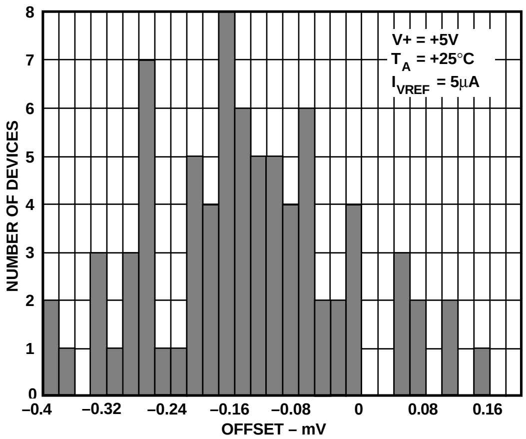


Figure 13. Comparator Input Offset Distribution

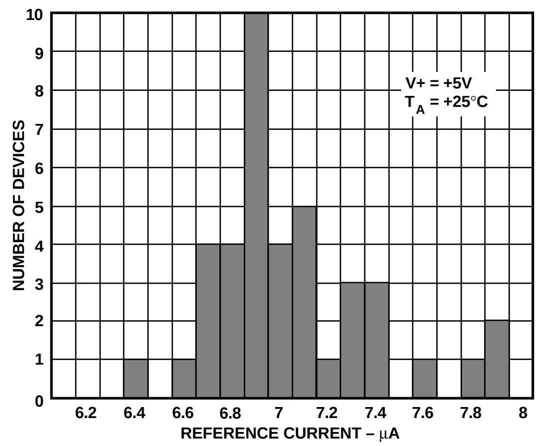


Figure 14. Zero Hysteresis Current Distribution

APPLICATIONS INFORMATION

Self-Heating Effects

In some applications the user should consider the effects of self-heating due to the power dissipated by the open-collector outputs, which are capable of sinking 20 mA continuously. Under full load, the TMP01 open-collector output device is dissipating

$$P_{DISS} = 0.6\ V \times .020A = 12\ mW$$

which in a surface-mount SO package accounts for a temperature increase due to self-heating of

$$\Delta T = P_{DISS} \times \theta_{JA} = .012\ W \times 158°C/W = 1.9°C.$$

This will of course directly affect the accuracy of the TMP01 and will for example cause the device to switch the heating output "OFF" 2 degrees early. Alternatively, bonding the same package to a moderate heatsink limits the self-heating effect to approximately

$$\Delta T = P_{DISS} \times \theta_{JC} = .012\ W \times 43°C/W = 0.52°C.$$

which is a much more tolerable error in most systems. The VREF and VPTAT outputs are also capable of delivering sufficient current to contribute heating effects and should not be ignored.

Buffering the Voltage Reference

As mentioned before, the reference output VREF is used to generate the temperature setpoint programming voltages for the TMP01 and also is used to determine the hysteresis temperature band by the reference load current I_{VREF}. The on-board output buffer amplifier is typically capable of 500 µA output drive into as much as 50 pF load (max). Exceeding this load will affect the accuracy of the reference voltage, could cause thermal sensing errors due to dissipation, and may induce oscillations. Selection of a low drift buffer functioning as a voltage follower with high input impedance will ensure optimal reference accuracy, and will not affect the programmed hysteresis current. Amplifiers which offer the low drift, low power consumption, and low cost appropriate to this application include the OP295, and members of the OP90, OP97, OP177 families, and others as shown in the following applications circuits.

With excellent drift and noise characteristics, VREF offers a good voltage reference for data acquisition and transducer excitation applications as well. Output drift is typically better than –10 ppm/°C, with 315 nV/$\sqrt{Hz}$ (typ) noise spectral density at 1 kHz.

Preserving Accuracy Over Wide Temperature Range Operation

The TMP01 is unique in offering both a wide-range temperature sensor and the associated detection circuitry needed to implement a complete thermostatic control function in one monolithic device. While the voltage reference, setpoint comparators, and output buffer amplifiers have been carefully compensated to maintain accuracy over the specified temperature range, the user has an additional task in maintaining the accuracy over wide operating temperature ranges in this application. Since the TMP01 is both sensor and control circuit, in many applications it is possible that the external components used to program and interface the device may be subjected to the same temperature extremes. Thus it may be necessary to locate components in close thermal proximity to minimize large temperature differentials, and to account for thermal drift errors where appropriate, such as resistor matching tempcos, amplifier error drift, and the like. Circuit design with the TMP01 requires a slightly different perspective regarding the thermal behavior of electronic components.

Thermal Response Time

The time required for a temperature sensor to settle to a specified accuracy is a function of the thermal mass of the sensor, and the thermal conductivity between the sensor and the object being sensed. Thermal mass is often considered equivalent to capacitance. Thermal conductivity is commonly specified using the symbol Q, and can be thought of as the reciprocal of thermal resistance. It is commonly specified in units of degrees per watt of power transferred across the thermal joint. Thus, the time required for the TMP01 to settle to the desired accuracy is dependent on the package selected, the thermal contact established in that particular application, and the equivalent power of the heat source. In most applications, the settling time is probably best determined empirically.

Switching Loads With The Open-Collector Outputs

In many temperature sensing and control applications some type of switching is required. Whether it be to turn on a heater when the temperature goes below a minimum value or to turn off a motor that is overheating, the open-collector outputs Over and Under can be used. For the majority of applications, the switches used need to handle large currents on the order of 1 amp and above. Because the TMP01 is accurately measuring temperature, the open-collector outputs should handle less than 20 mA of current to minimize self-heating. Clearly, the Over-temp and Under-temp outputs should not drive the equipment directly. Instead, an external switching device is required to handle the large currents. Some examples of these are relays, power MOSFETs, thyristors, IGBTs, and Darlingtons.

Figure 15 shows a variety of circuits where the TMP01 controls a switch. The main consideration in these circuits, such as the relay in Figure 15a, is the current required to activate the switch.

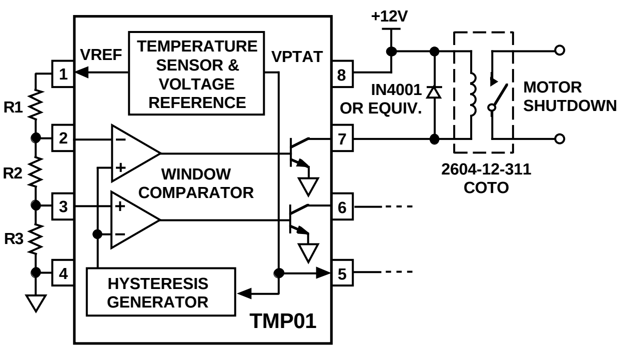


Figure 15a. Reed Relay Drive

It is important to check the particular relay you choose to ensure that the current needed to activate the coil does not exceed the TMP01's recommended output current of 20 mA. This is easily determined by dividing the relay coil voltage by the specified coil resistance. Keep in mind that the inductance of the relay will create large voltage spikes that can damage the TMP01 output unless protected by a commutation diode across the coil, as shown. The relay shown has a contact rating of 10 watts maximum. If a relay capable of handling more power is desired, the larger contacts will probably require a commensurately larger coil, with lower coil resistance and thus higher trigger current. As the contact power handling capability increases, so does the current needed for the coil. In some cases an external driving transistor should be used to remove the current load on the TMP01 as explained in the next section.

Power FETs are popular for handling a variety of high current DC loads. Figure 15b shows the TMP01 driving a p-channel MOSFET transistor for a simple heater circuit. When the output transistor turns on, the gate of the MOSFET is pulled down to approximately 0.6 V, turning it on. For most MOSFETs a gate-to-source voltage or Vgs on the order of –2 V to –5 V is sufficient to turn the device on. Figure 15c shows a similar circuit for turning on an n-channel MOSFET, except that now the gate to source voltage is positive. Because of this reason an external transistor must be used as an inverter so that the MOSFET will turn on when the "Under Temp" output pulls down.

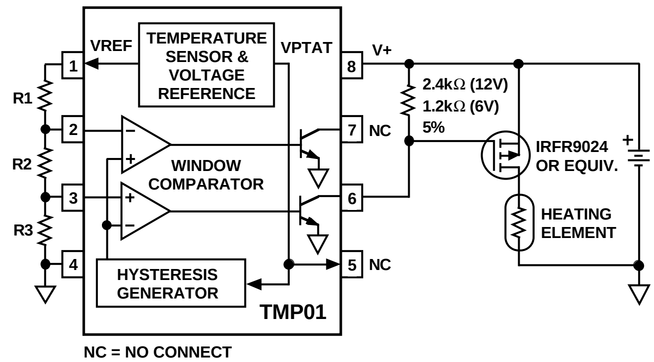


Figure 15b. Driving a P-Channel MOSFET

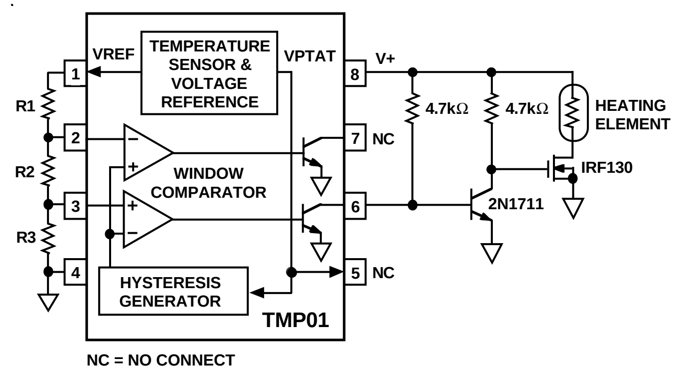


Figure 15c. Driving a N-Channel MOSFET

Isolated Gate Bipolar Transistors (IGBT) combine many of the benefits of power MOSFETs with bipolar transistors, and are used for a variety of high power applications. Because IGBTs have a gate similar to MOSFETs, turning on and off the devices is relatively simple as shown in Figure 15d. The turn on voltage for the IGBT shown (IRGBC40S) is between 3.0 and 5.5 volts. This part has a continuous collector current rating of 50 A and a maximum collector to emitter voltage of 600 V, enabling it to work in very demanding applications.

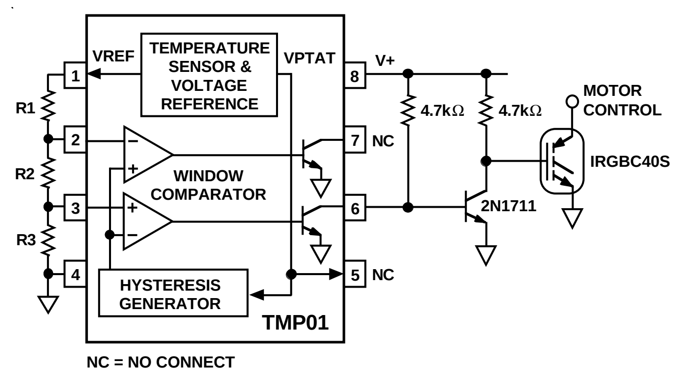


Figure 15d. Driving an IGBT

The last class of high power devices discussed here are Thyristors, which includes SCRs and Triacs. Triacs are a useful alternative to relays for switching ac line voltages. The 2N6073A shown in Figure 15e is rated to handle 4A (rms). The optoisolated MOC3011. Triac shown features excellent electrical isolation from the noisy ac line and complete control over the high power Triac with only a few additional components.

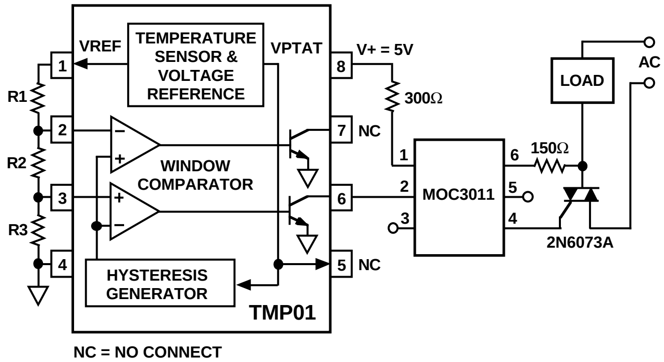


Figure 15e. Controlling the 2N6073A Triac

High Current Switching

As mentioned above, internal dissipation due to large loads on the TMP01 outputs will cause some temperature error due to self-heating. External transistors remove the load from the TMP01, so that virtually no power is dissipated in the internal transistors and no self-heating occurs. Figure 16 shows a few examples using external transistors. The simplest case, using a single transistor on the output to invert the output signal is shown in Figure 16a. When the open-collector of the TMP01 turns "ON" and pulls the output down, the external transistor Q1's base will be pulled low, turning off the transistor. Another transistor can be added to reinvert the signal as shown in Figure 16b. Now, when the output of the TMP01 is pulled down, the first transistor, Q1, turns off and its collector goes high, which turns Q2 on, pulling its collector low. Thus, the output taken from the collector of Q2 is identical to the output of the TMP01. By picking a transistor that can accommodate large amounts of current, many high power devices can be switched.

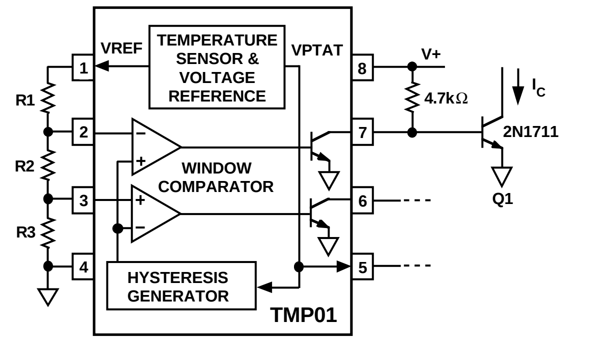


Figure 16a. An External Resistor Minimizes Self-Heating

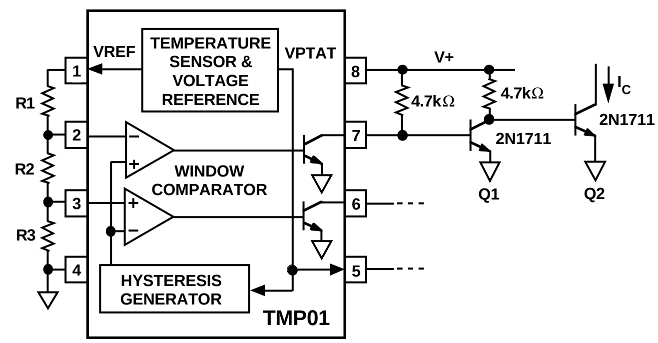


Figure 16b. Second Transistor Maintains Polarity of TMP01 Output

An example of a higher power transistor is a standard Darlington configuration as shown in Figure 16c. The part chosen, TIP-110, can handle 2A continuous which is more than enough to control many high power relays. In fact the Darlington itself can be used as the switch, similar to MOSFETs and IGBTs.

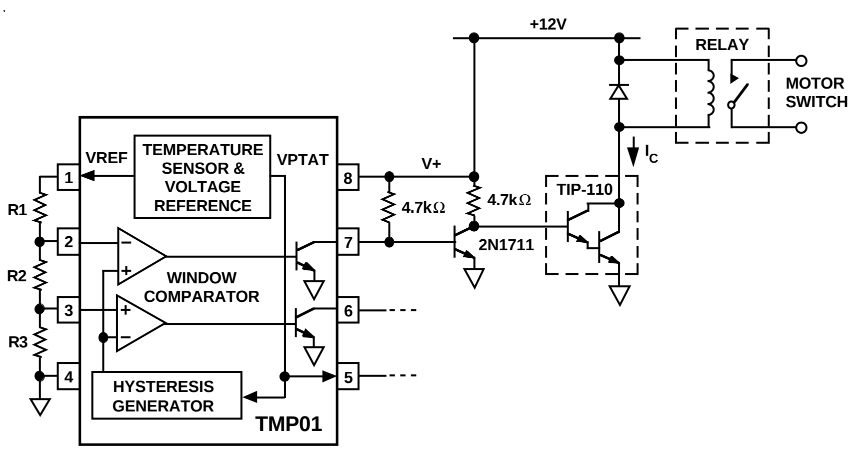


Figure 16c. Darlington Transistor Can Handle Large Currents

Buffering the Temperature Output Pin

The VPTAT sensor output is a low impedance dc output voltage with a 5 mV/K temperature coefficient, and is useful in a number of measurement and control applications. In many applications, this voltage needs to be transmitted to a central location for processing. The buffered VPTAT voltage output is capable of 500 μA drive into 50 pF (max). As mentioned in the discussion above regarding buffering circuits for the VREF output, it is useful to consider external amplifiers for interfacing VPTAT to external circuitry to ensure accuracy, and to minimize loading which could create dissipation-induced temperature sensing errors. An excellent general-purpose buffer circuit using the OP177 is shown in Figure 17 which is capable of driving over 10 mA, and will remain stable under capacitive loads of up to 0.1 μF. Other interfacing ideas are shown below.

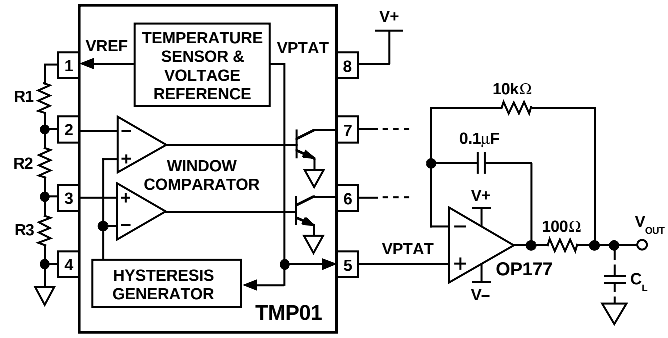


Figure 17. Buffer VPTAT to Handle Difficult Loads

Differential Transmitter

In noisy industrial environments, it is difficult to send an accurate analog signal over a significant distance. However, by sending the signal differentially on a wire pair, these errors can be significantly reduced. Since the noise will be picked up equally on both wires, a receiver with high common-mode input rejection can be used to cancel out the noise very effectively at the receiving end. Figure 18 shows two amplifiers being used to send the signal differentially, and an excellent differential receiver, the AMP03, which features a common-mode rejection ratio of 95 dB at dc and very low input and drift errors.

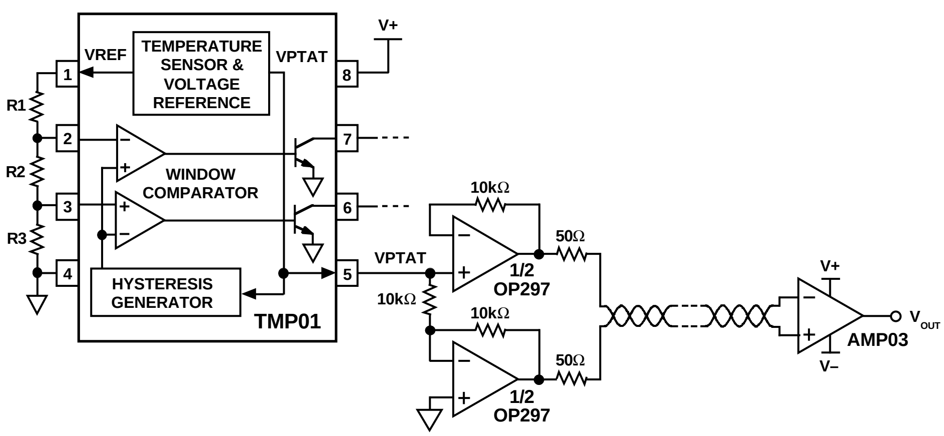


Figure 18. Send the Signal Differentially for Noise Immunity

4 mA-20 mA Current Loop

Another, very common method of transmitting a signal over long distances is to use a 4 mA-20 mA Loop, as shown in Figure 19. An advantage of using a 4 mA-20 mA loop is that the accuracy of a current loop is not compromised by voltage drops across the line. One requirement of 4 mA-20 mA circuits is that the remote end must receive all of its power from the loop, meaning that the circuit must consume less than 4 mA. Operating from +5 V, the quiescent current of the TMP01 is 500 μA max, and the OP90s is 20 μA max, totaling less than 4 mA. Although not shown, the open collector outputs and temperature setting pins can be connected to do any local control of switching.

The current is proportional to the voltage on the VPTAT output, and is calibrated to 4 mA at a temperature of –40°C, to 20 mA for +85°C. The main equation governing the operation of this circuit gives the current as a function of VPTAT:

$$I_{OUT} = \frac{1}{R6}\left(\frac{VPTAT \times R5}{R2} - \frac{VREF \times R3}{R3 + R1}\left(1 + \frac{R5}{R2}\right)\right)$$

The resulting temperature coefficient of the output current is 128 μA/°C.

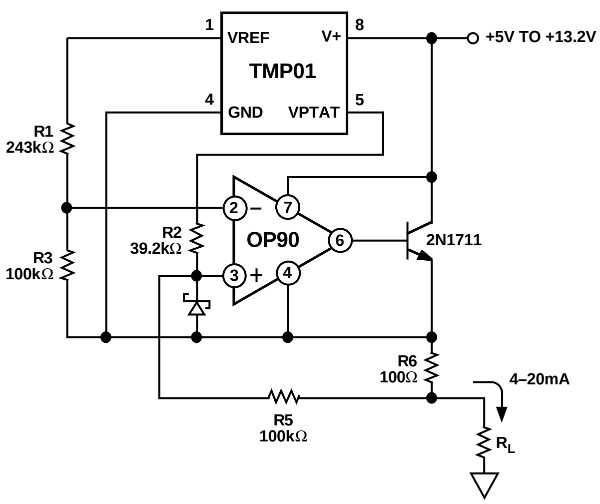


Figure 19. 4-20 mA Current Loop

To determine the resistor values in this circuit, first note that VREF remains constant over temperature. Thus the ratio of R5 over R2 must give a variation of I_{OUT} from 4 mA to 20 mA as VPTAT varies from 1.165 V at –40°C to 1.79 V at +85°C. The absolute value of the resistors is not important, only the ratio. For convenience, 100 kΩ is chosen for R5. Once R2 is calculated, the value of R3 and R1 is determined by substituting 4 mA for I_{OUT} and 1.165 V for VPTAT and solving. The final values are shown in the circuit. The OP90 is chosen for this circuit because of its ability to operate on a single supply and its high accuracy. For initial accuracy, a 10 kΩ trim potentiometer can be included in series with R3, and the value of R3 lowered to 95 kΩ. The potentiometer should be adjusted to produce an output current of 12.3 mA at 25°C.

Temperature-to-Frequency Converter

Another common method of transmitting analog information is to convert a voltage to the frequency domain. This is easily done with any of the low cost monolithic Voltage-to-Frequency Converters (VFCs) available, which feature a robust, open-collector digital output. A digital signal is very immune to noise and voltage drops because the only important information is the frequency. As long as the conversions between temperature and frequency are done accurately, the temperature data can be successfully transmitted.

A simple circuit to do this combines the TMP01 with an AD654 VFC, as shown in Figure 20. The AD654 outputs a square wave that is proportional to the dc input voltage according to the following equation:

$$F_{OUT} = \frac{V_{IN}}{10\,(R1 + R2)\,C_T}$$

By simply connecting the VPTAT output to the input of the AD654, the 5 mV/°C temperature coefficient gives a sensitivity of 25 Hz/°C, centered around 7.5 kHz at 25°C. The trimming resistor R2 is needed to calibrate the absolute accuracy of the AD654. For more information on that part, please consult the AD654 data sheet. Finally, the AD650 can be used to accurately convert the frequency back to a dc voltage on the receiving end.

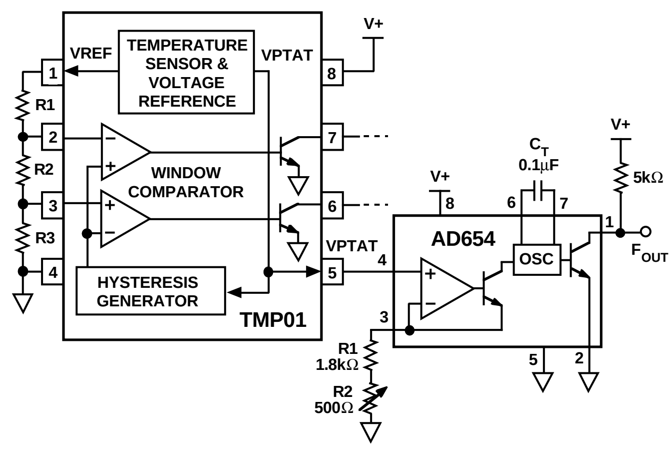


Figure 20. Temperature-to-Frequency Converter

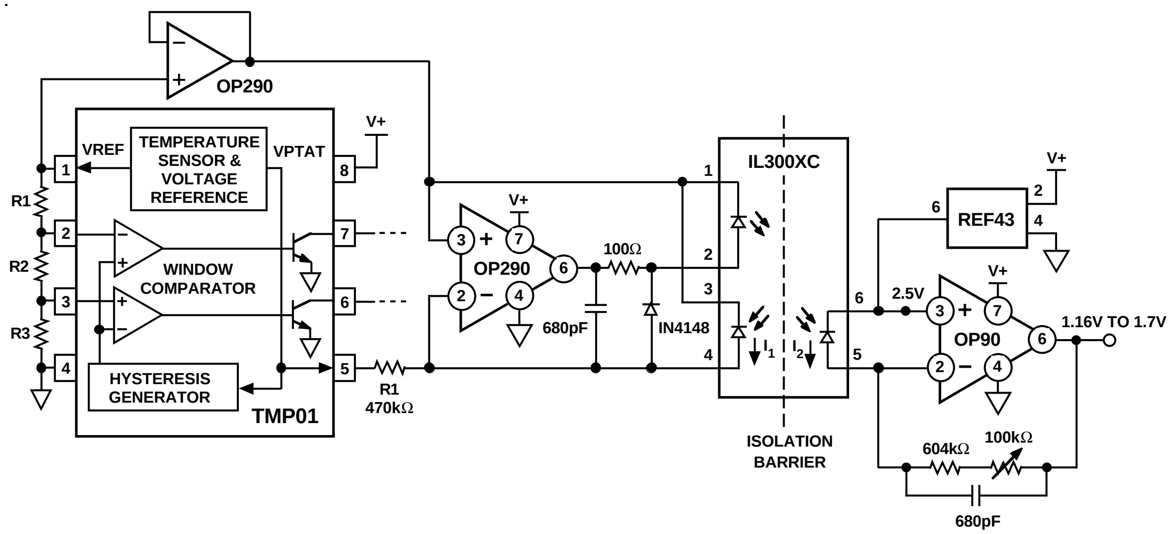


Figure 21. Isolation Amplifier

Isolation Amplifier

In many industrial applications the sensor is located in an environment that needs to be electrically isolated from the central processing area. Figure 21 shows a simple circuit that uses an 8-pin optoisolator (IL300XC) that can operate across a 5,000 V barrier. IC1 (an OP290 single-supply amplifier) is used to drive the LED connected between Pins 1 to 2. The feedback actually comes from the photodiode connected from Pins 3 to 4. The OP290 drives the LED such that there is enough current generated in the photodiode to exactly equal the current derived from the VPTAT voltage across the 470 kΩ resistor. On the receiving end, an OP90 converts the current from the second photodiode to a voltage through its feedback resistor R2. Note that the other amplifier in the dual OP290 is used to buffer the 2.5 V reference voltage of the TMP01 for an accurate, low drift LED bias level without affecting the programmed hysteresis current. A REF43 (a precision 2.5 V reference) provides an accurate bias level at the receiving end.

To understand this circuit, it helps to examine the overall equation for the output voltage. First, the current (I1) in the photodiode is set by:

$$I_1 = \frac{2.5\,V - VPTAT}{470\ k\Omega}$$

Note that the IL300XC has a gain of 0.73 (typical) with a min and max of 0.693 and 0.769 respectively. Since this is less than 1.0, R2 must be larger than R1 to achieve overall unity gain. To show this the full equation is:

$$V_{OUT} = 2.5\,V - I_2 R_2 = 2.5\,V - 0.7\left(\frac{2.5\,V - VPTAT}{470\ k\Omega}\right) 644\ k\Omega = VPTAT$$

A trim is included for R2 to correct for the initial gain accuracy of the IL300XC. To perform this trim, simply adjust for an output voltage equal to VPTAT at any particular temperature. For example, at room temperature, VPTAT = 1.49 V, so adjust R2 until V_{OUT} = 1.49 V as well. Both the REF43 and the OP90 operate from a single supply, and contribute no significant error due to drift.

In order to avoid the accuracy trim, and to reduce board space, complete isolation amplifiers are available, such as the high accuracy AD202.

Out-of-Range Warning

By connecting the two open collector outputs of the TMP01 together into a "wired-OR" configuration, a temperature "out-of-range" warning signal is generated. This can be useful in sensitive equipment calibrated to work over a limited temperature range. R1, R2, and R3 in Figure 22 are chosen to give a temperature range of 10°C around room temperature (25°C). Thus, if the temperature in the equipment falls below +15°C or rises above +35°C, the Undertemp Output or Overtemp Output respectively will go low and turn the LED on. The LED may be replaced with a simple pull-up resistor to give a logic output for controlling the instrument, or any of the switching devices discussed above can be used.

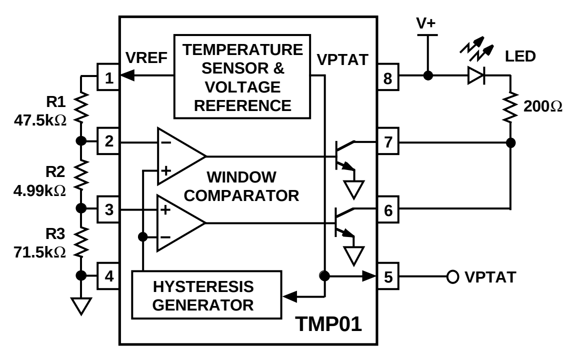


Figure 22. Out-of-Range Warning

Translating 5 mV/K to 10 mV/°C

A useful circuit is shown in Figure 23 that translates the VPTAT output voltage, which is calibrated in Kelvins, into an output that can be read directly in degrees Celsius on a voltmeter display. To accomplish this, an external amplifier is configured as a differential amplifier. The resistors are scaled so the VREF voltage will exactly cancel the VPTAT voltage at 0.0°C.

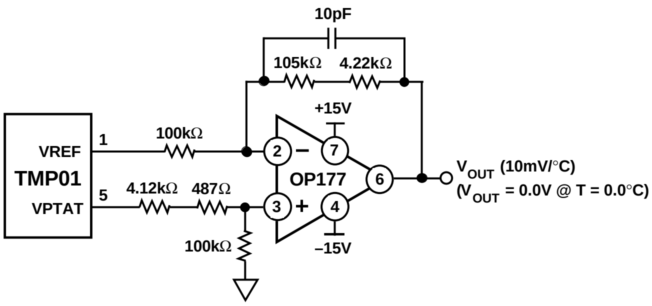


Figure 23. Translating 5 mV/K to 10 mV/°C

However, the gain from VPTAT to the output is two, so that 5 mV/K becomes 10 mV/°C. Thus, for a temperature of +80°C, the output voltage is 800 mV. Circuit errors will be due primarily to the inaccuracies of the resistor values. Using 1% resistors the observed error was less than 10 mV, or 1°C. The 10 pF feedback capacitor helps to ensure against oscillations. For better accuracy, a adjustment potentiometer can be added in series with either 100 kΩ resistor.

Translating VPTAT to the Fahrenheit Scale

A very similar circuit to the one shown in Figure 23 can be used to translate VPTAT into an output that can be read directly in degrees Fahrenheit, with a scaling of 10 mV/°F. Only unity gain or less is available from the first stage differentiating circuit, so the second amplifier provides a gain of two to complete the conversion to the Fahrenheit scale. Using the circuit in Figure 24, a temperature of 0.0°F gives an output of 0.00 V. At room temperature (70°F) the output voltage is 700 mV. A –40°C to +85°C operating range translates into –40°F to +185°F. The errors are essentially the same as for the circuit in Figure 23.

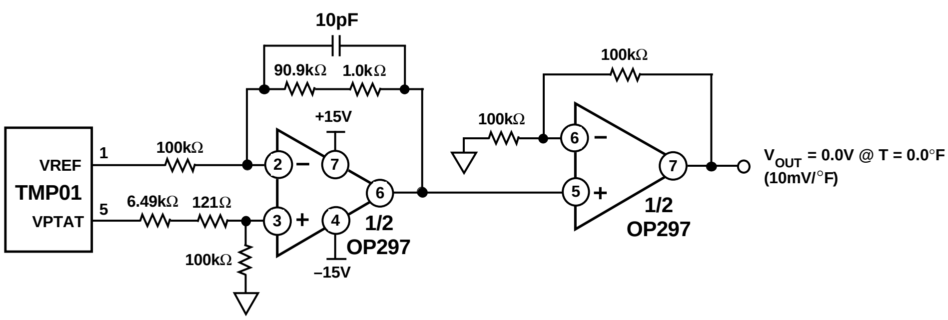
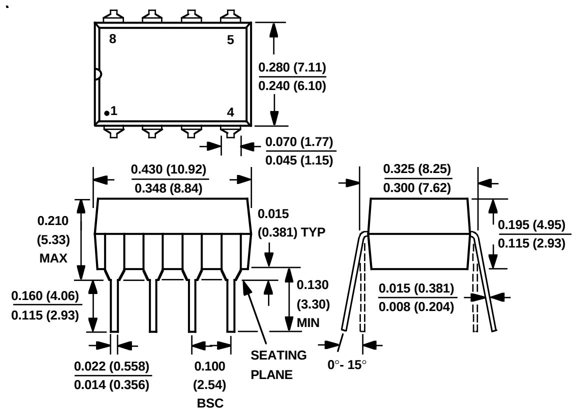


Figure 24. Translating 5 mV/K to 10 mV/°F

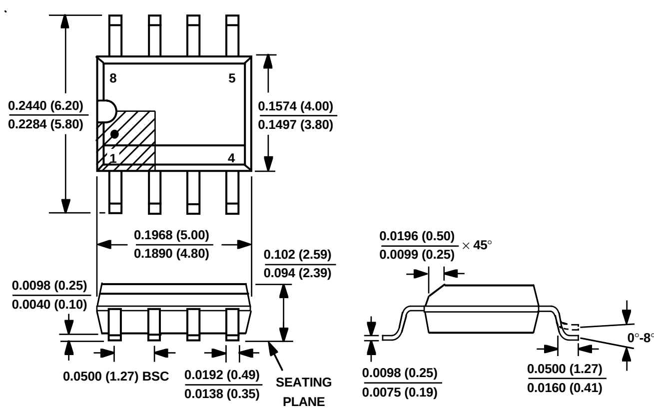
OUTLINE DIMENSIONS

Dimensions shown in inches and (mm).

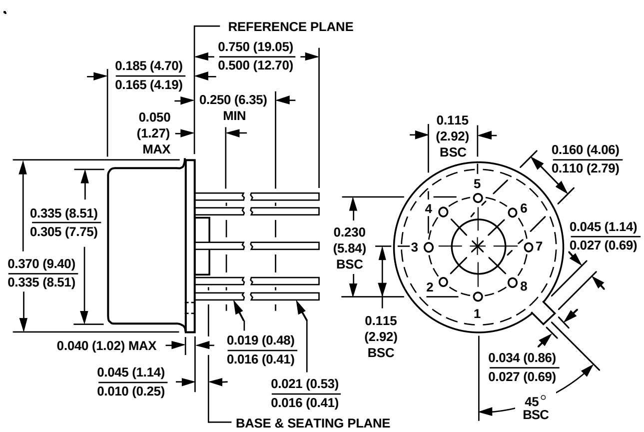
8-Pin Epoxy DIP



8-Pin SOIC



8-Pin TO-99



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